TRANSISTORIZED VARIABLE DUTY CYCLE D. C. VOLTAGE REGULATOR

RICHARD M. HARVEY

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by

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Submitted in partial fulfillment of
the requirements for the degree of

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ABSTRACT

The specifications of a missile's power supply require that a variation of output voltage of within .1% be maintained from an unregulated input voltage varying as much as 20% from its quiescent value, output load variation of .1 to .4 amperes, and an efficiency equal to or better than a series regulator.

This is a report of experimental and analytical investigations of the factors influencing the efficiency of a variable duty cycle transistorized regulator designed to meet these specifications. Basic to the theoretical efficiency of this type of regulator is the relation:

$$\text{duty cycle} = \frac{\text{conduction time}}{\text{total period}}$$

where "conduction time" is the interval during which current is passed through a low loss inductor in series between the source and load. The load voltage, before final filtering, controls the conduction time through appropriate feed-back circuitry. A discussion of the required and acceptable circuitry and a summary of the characteristics of the regulator is included.

The writer wishes to express his appreciation for the assistance and encouragement given him by Mr. E. Mueller of the Lockheed Missile and Space Company in this investigation.

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ILLUSTRATIONS

INTRODUCTION

Modulized missile electronics circuits require a high efficiency voltage regulator that is capable of providing an output voltage constantly maintained within one percent of a specified value under variable input voltage, output load, and ambient temperature conditions. A series regulator can be designed to meet these specifications, but such a series regulator must dissipate all excess power during periods of high input voltage and load, and is thus limited in its theoretical maximum efficiency. Heat generated by this power dissipation creates problems in module design, where heat sinks and temperature sensitive components often cannot be physically isolated in the same module.

Present series regulators are limited in efficiency by the relation

$$\text{Efficiency} = \frac{V_{out} \times I_{out}}{V_{in} \times I_{in}} .$$

If current in and current out are the same, as in an ideal series regulator, then efficiency is a function of voltage only,

$$\text{Efficiency} = \frac{V_{out}}{V_{in}}$$

and all voltage above the regulated output is dissipated as heat. Power dissipation may be given by the relation

$$P_{Diss.} = -\left(V_{out} - V_{in}\right) I_{out} .$$

If a fluctuating input voltage is provided to be regulated, and this input voltage consistently runs 50% over the regulated output voltage with only occasional drops to near the regulated level, a large amount of power is dissipated most of the time, as shown in Figure (1).

The particualr duty cycle regulator employed in this analysis was designed to meet the following specifications:

Input Voltage:	24 - 36 Volts
Output Voltage:	22.000 Volts
Output Voltage Variation:	.1%
Load Variation:	.1 -.4 amps
Efficiency:	85% or greater

Output voltage is preset at the desired value, but the system can be modified to provide any output voltage above 10 volts depending on components and non-regulated source voltage available. A comparison of efficiency of an ideal series voltage regulator and the variable duty cycle regulator analyzed is shown in Figure (1).

Transistor circuitry is employed. Applications in mobile units where a limited amount of power is available, printed circuit heat dissipation is a problem, and good regulation is mandatory for precision switching or amplification circuits, justify the more complex and expensive duty cycle regulator.

A general discussion of the overall circuit is presented to acquaint the reader with the interaction of each section and provide a general basis for further analysis. Each section is then delineated with respect to its function in the circuit, and is studied independently. Waveforms, component variations, circuit substitutions, and design considerations are presented in the discussion of the individual sections. Conclusions and comments are made in hopes that this circuit may be beneficially used in other applications.

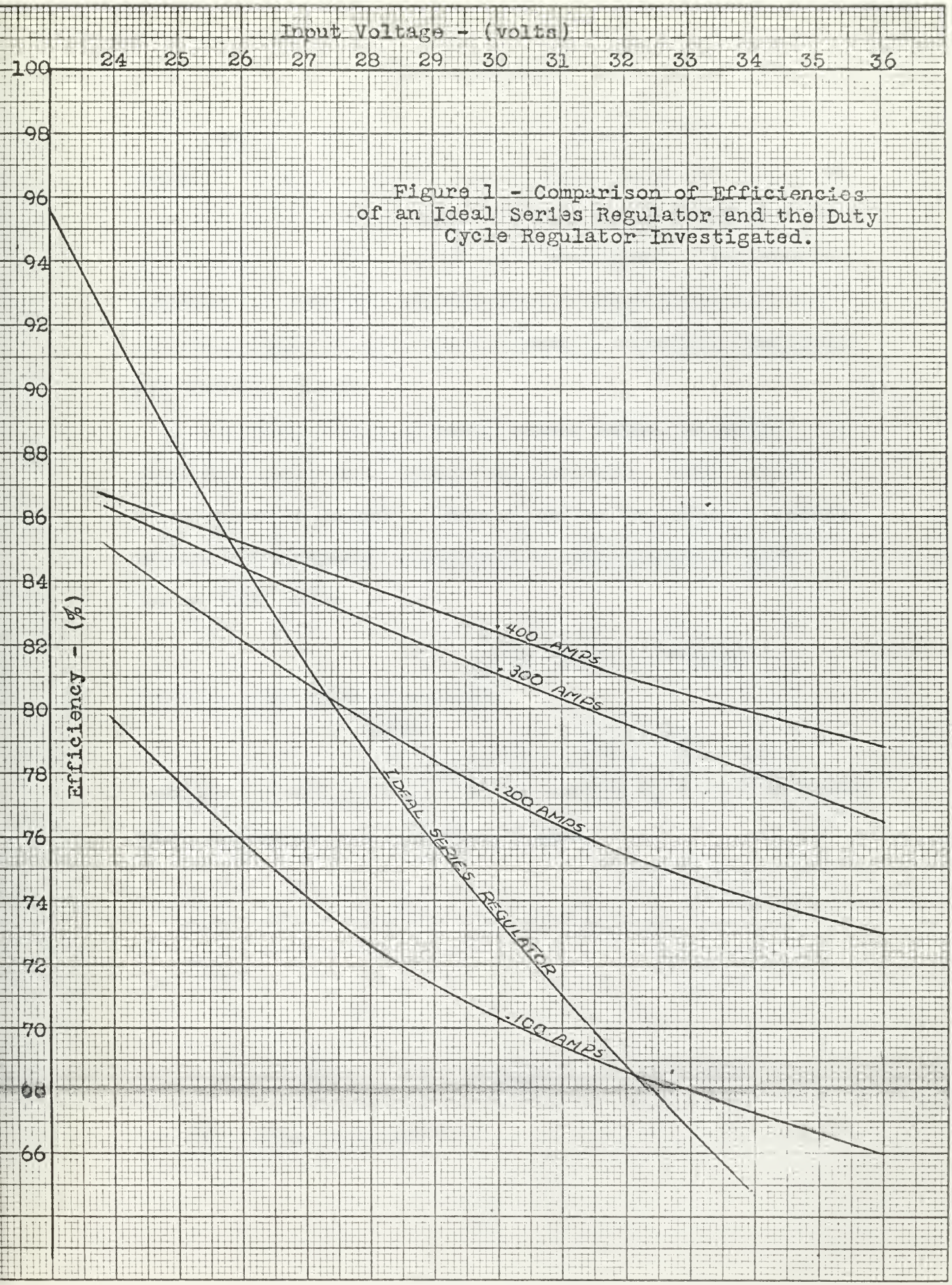


Figure 1 - Comparison of Efficiencies of an Ideal Series Regulator and the Duty Cycle Regulator Investigated.

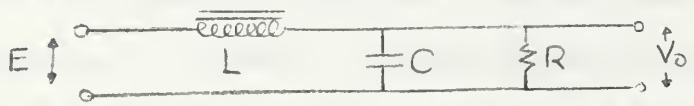
GENERAL OPERATION

A magnetic field contains potential energy. Electrical energy is changed to magnetic flux when building up the field of an iron core inductor, and changed back to electrical energy when the field collapses. Efficiency of energy transformation is essentially determined by hysteresis and eddy current losses in the core, and if both are small the core may be likened to a reservoir that stores and discharges energy.

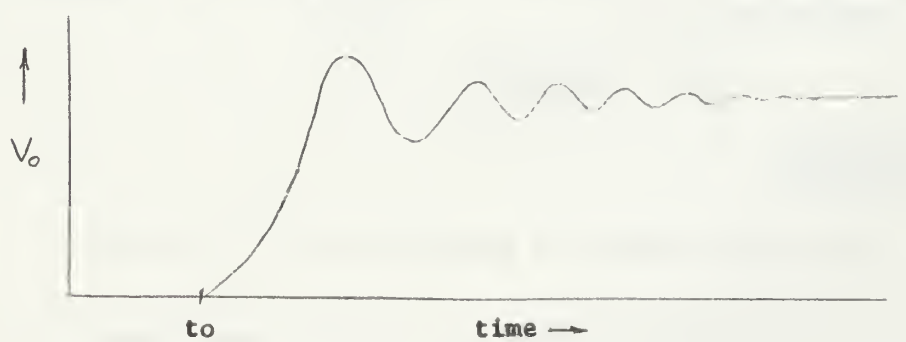
A capacitor is also an energy storage device, but stores electrical energy itself without any transition. Cogent use of these properties provides a means of storing the energy of a square DC wave and essentially filtering it to some integral DC value. For an ideal filter,

$$\frac{1}{T}\int_{0}^{T} E_{in}\, I_{in}\, dt = E_{out}\, I_{out}$$

For a low pass filter of the type

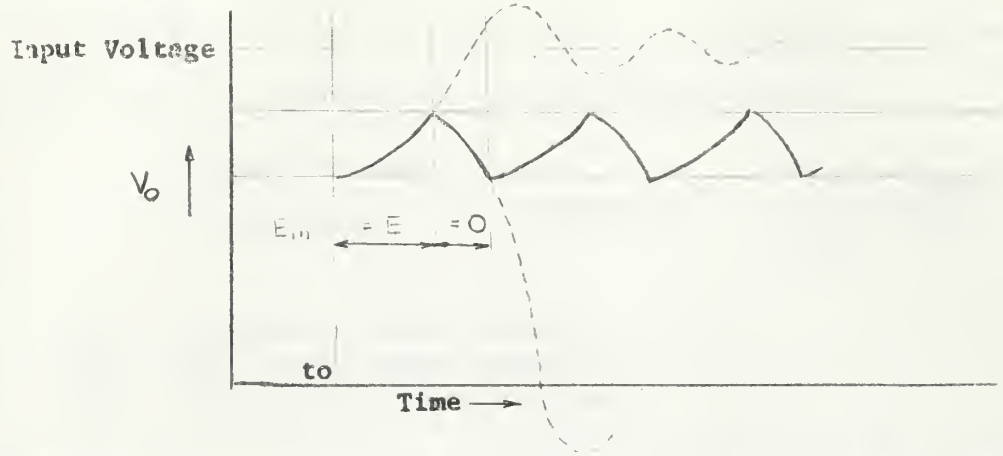


instantaneous voltage is given by the equation $V_o = E\,(i - e^{-\alpha t} \sin(\omega t + \phi))$ * which in response to a step input is shown.



*Derivation in Appendix 3

If it is assumed that the time constant of the filter is much much greater than the period of the input square wave, then the filtered steady state ripple voltage will appear in the following manner:



By varying the conduction time of the input square wave at constant current output, some greater integral of output voltage is achieved, so that the output voltage of the filter may be controlled by lengthening or decreasing the conduction time per cycle.

This is the theory of operation of the variable duty cycle voltage regulator, a schematic of which appears in Figure (2), and block diagram in Figure (3).

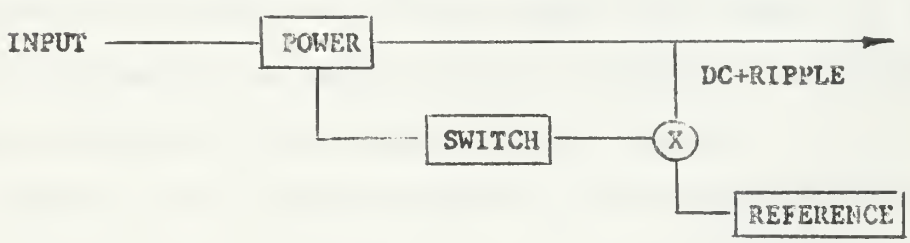


Figure (3) - Block diagram of variable duty cycle regulator.

At steady state operation, a switching circuit is excited by the extremes of the ripple voltage and varies the width of the duty cycle to maintain a constant DC voltage output.

As shown in the block diagram the regulator can be functionally divided into the voltage reference circuit, the switching circuit, and the power conduction circuit. These subcircuits are discussed individually.

Frequency and duty cycle variation are independent. Pulse width is ideally determined by the following relation:

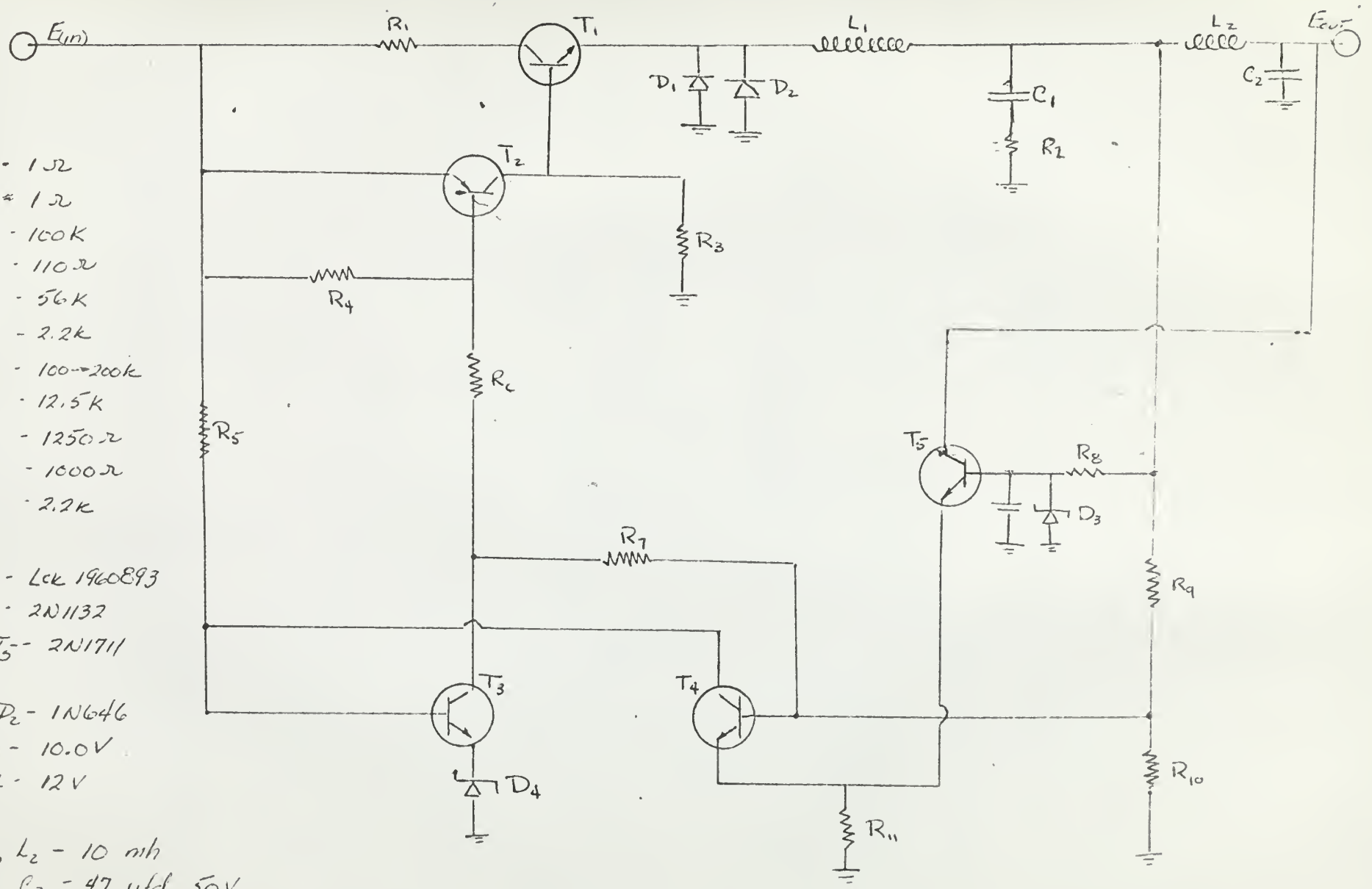
$$\frac{\text{Conduction Time}}{\text{Total period}} = \text{Duty cycle} = \frac{V_{out} + V_{drop}}{V_{in}}$$

V_{out} - Regulated output voltage
V_{drop} - Drop across power transistor
V_{in} - Nonregulated input voltage

Frequency is determined by a combination of the magnitude of the ripple voltage and the time constant of the inductance and capacitance of the filter, and the output current.

High frequency noise caused by sharp current spikes and hard switching action requires short connections and shielding of the entire unit from other electronic components. Output ripple, dependent on the final stage filtering, is of the order of .001%. Temperature sensitivity is .07% per degree in the range -25° to $+150^{\circ}$ Centigrade, but may be reduced to .007% per degree Centigrade by the insulation of the voltage reference component.

Figure (2) is a complete schematic of the voltage regulator.



R_1 - 1 Ω
R_2 - 1 Ω
R_3 - 100K
R_4 - 110 Ω
R_5 - 56K
R_6 - 2.2K
R_7 - 100→200K
R_8 - 12.5K
R_9 - 1250 Ω
R_{10} - 1000 Ω
R_{11} - 2.2K

T_1 - Lck 1960893
T_2 - 2N1132
T_3 - T_5 - 2N1711

D_1 D_2 - 1N646
D_3 - 10.0V
D_4 - 12V

L_1, L_2 - 10 mh
C_1, C_2 - 47 µfd. 50V.

FIGURE (2) - BASIC SCHEMATIC

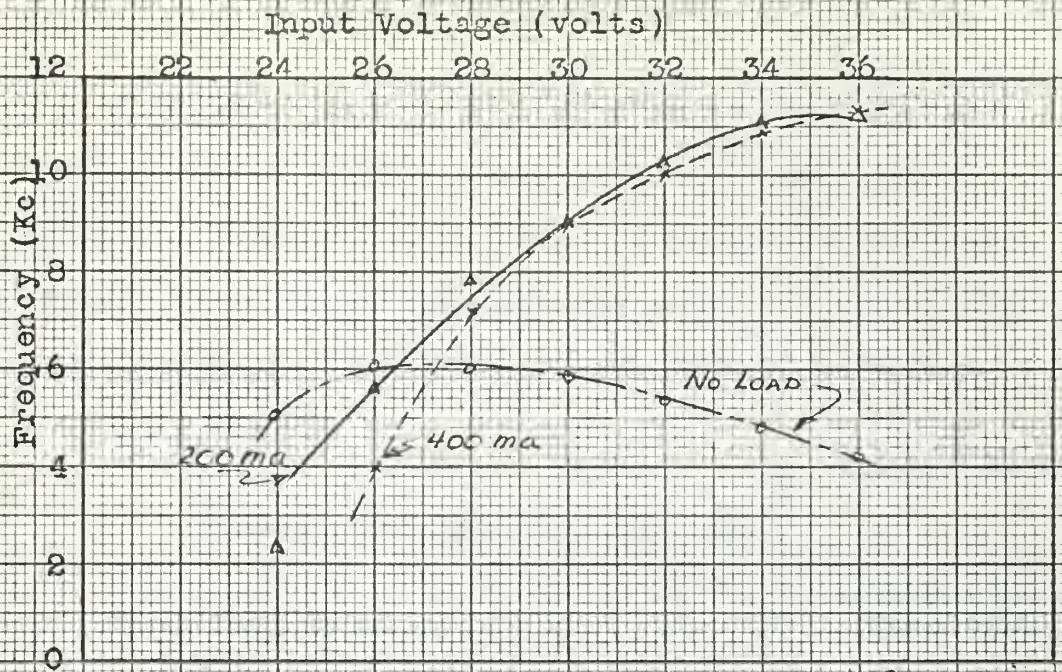


Figure 4(a) - Frequency vs Input Voltage at constant Load

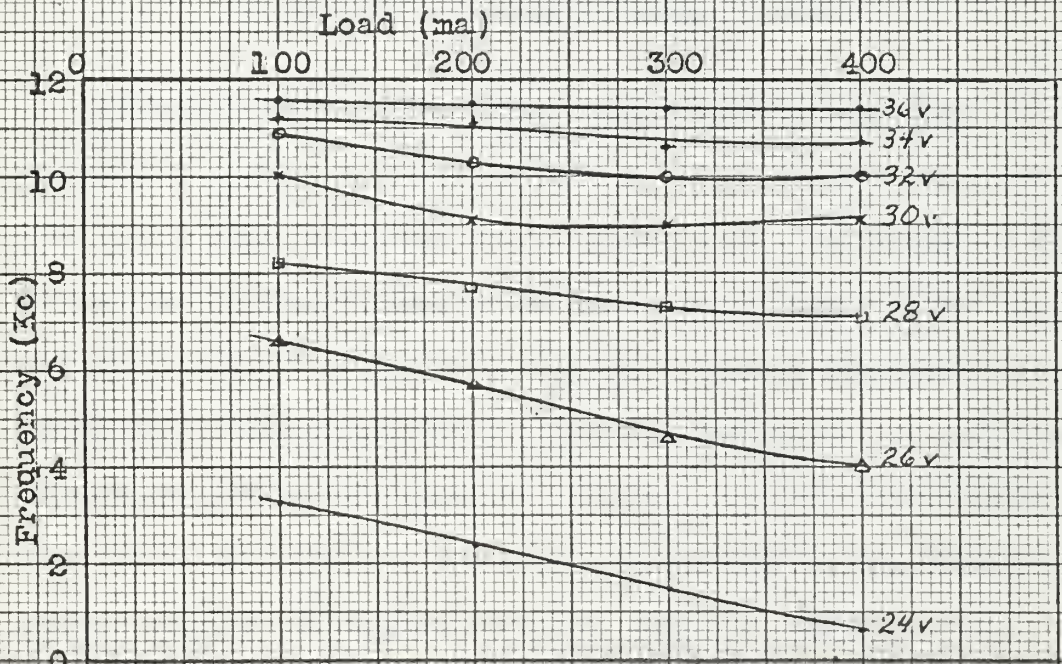


Figure 4(b)- Frequency vs Load at constant Input Voltage

VOLTAGE REFERENCE CIRCUIT

An amplified 'hard' reference voltage is provided by the voltage reference stage to ensure predictable and reliable switching. Regulator output voltage is compared to this hard voltage to initiate and stop switching action as load and input conditions demand. Figure (5) is a schematic of the voltage reference circuit.

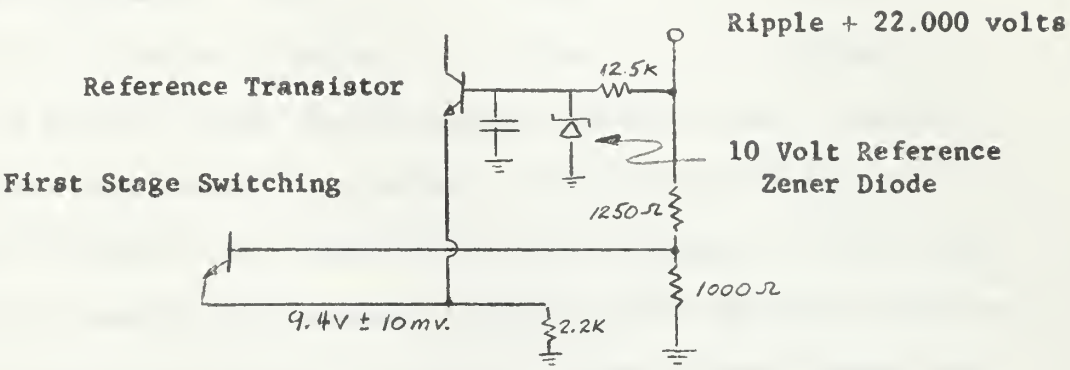


Figure (5) - Voltage Reference Circuit

Emitter voltage on the first stage switching transistor is maintained at about 9.4 volts within $\pm$ 10 mv. by the amplified effect of a high resistance 10 volt zener diode (TRANSITRON SV 4010A). Because the reference stage is operated in continuous saturation, emmitter potential of the switching and reference transistors is maintained constant regardless of current changes due to switching through the common 2.2K emmitter resistor. Current amplification required of the reference transistor is governed by the ratio of current required to maintain constant emmitter potential (i.e., current differential equal to that produced by first switching stage between switch on and switch off conditions) to allowable

current available through 12.5K resistor that will still allow zener operation on the knee of the characteristic curve. Table (1) is shown below comparing the regulator characteristics using lower beta transistors in the reference voltage circuit.

Quantity	B - 100	B - 45
Efficiency (400 ma)	87.4%	79.%
Load regulation 30V	10 mv	5 mv
Input regulation @ 400 ma	15 mv	92 mv

It was also noted that the voltage divider circuit supplying the first stage switching transistor (resistors R_9 and R_{10} of Figure (2)) had to be trimmed for an output voltage of 22.000 volts using the lower beta transistor, indicating that the base of the low beta reference transistor was pulling a higher current through the resistor R_8, and was thus operating lower on the knee of the reference zener diode, as shown in Figure (6) below. Operation without trimming the voltage divider circuit gave a regulator output voltage of 21.88 Volts.

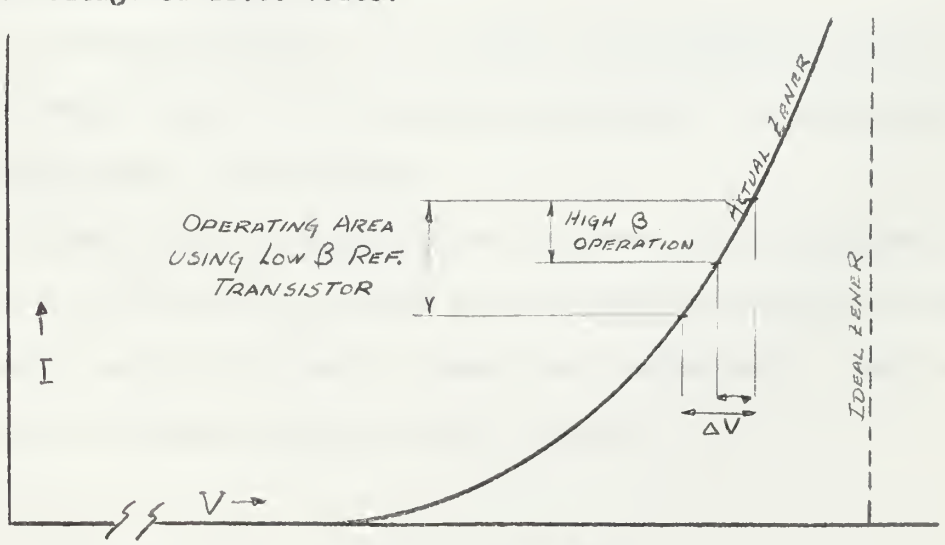


Figure (6) - Zener Diode operation

Operation of the voltage reference circuit using a more nearly ideal reference zener diode and low Beta transistor would have been more acceptable in that current surges of the voltage reference transistor would not have produced as large a change in base reference voltage on the characteristic curve (Figure 6). Unfortunately, an ideal zener diode does not exist, and those low resistance zener diodes that are more desirable in this respect have larger temperature coefficients (unless purchased as matched pairs, an expensive possibility) and thus render the Reference Circuit, and the regulator itself, unacceptably sensitive to ambient temperature. Effects of temperature on reference zener voltage for the Transitron 1040A 10 Volt zener diode is shown in Figure 7a. Output voltage as a function of zener reference voltage is shown in Figure 7b, and the combination is shown in Figure 8. Temperature sensitivity of the zener diode is amplified as shown in Figure 8, so that temperature sensitivity of the over all regulator is determined predominantly by the reference zener diode.

A 1 ufd capacitor and 12.5K resistor are necessary to keep the switching signal off the reference stage, and work the zener at its rated current, respectively.

Alternate circuit configurations are shown in the Appendix 2, pages 1-3, but were not used due to the high cost of precision diodes with sharp knees and allowable temperature coefficients. Note the considerable saving of parts in these circuits.

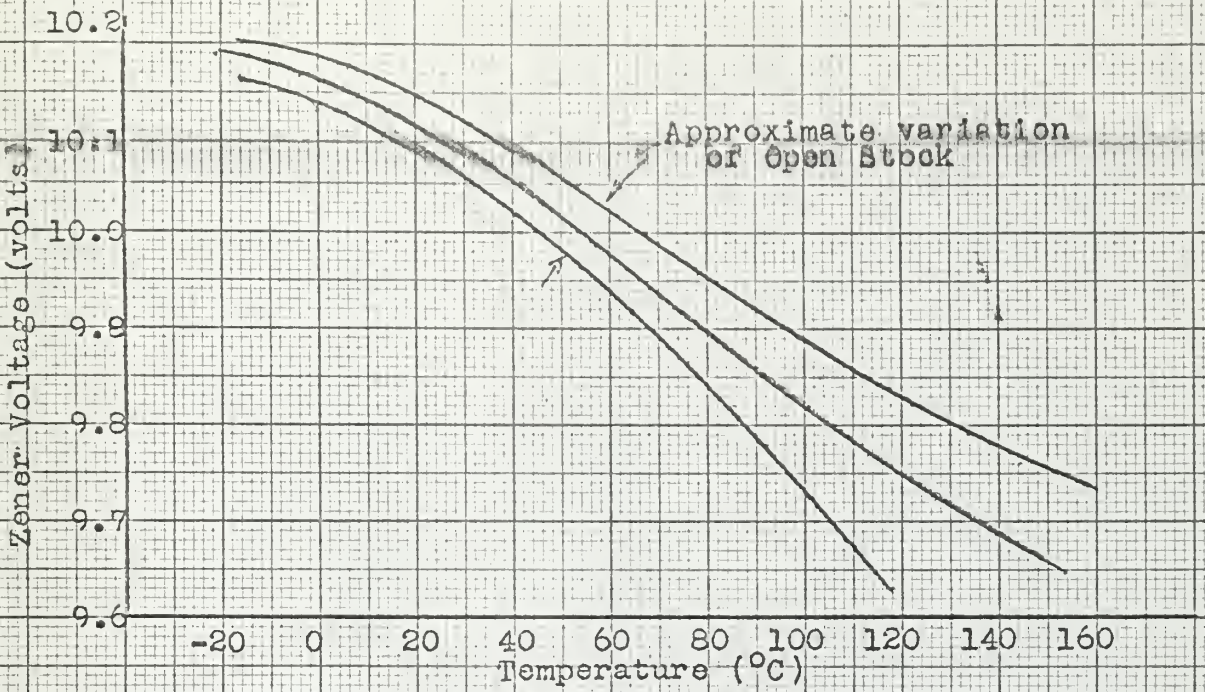


Figure 7(a) - Zener Voltage Variation with Temperature (@ 1 ma) - TRANSITRON 4010A

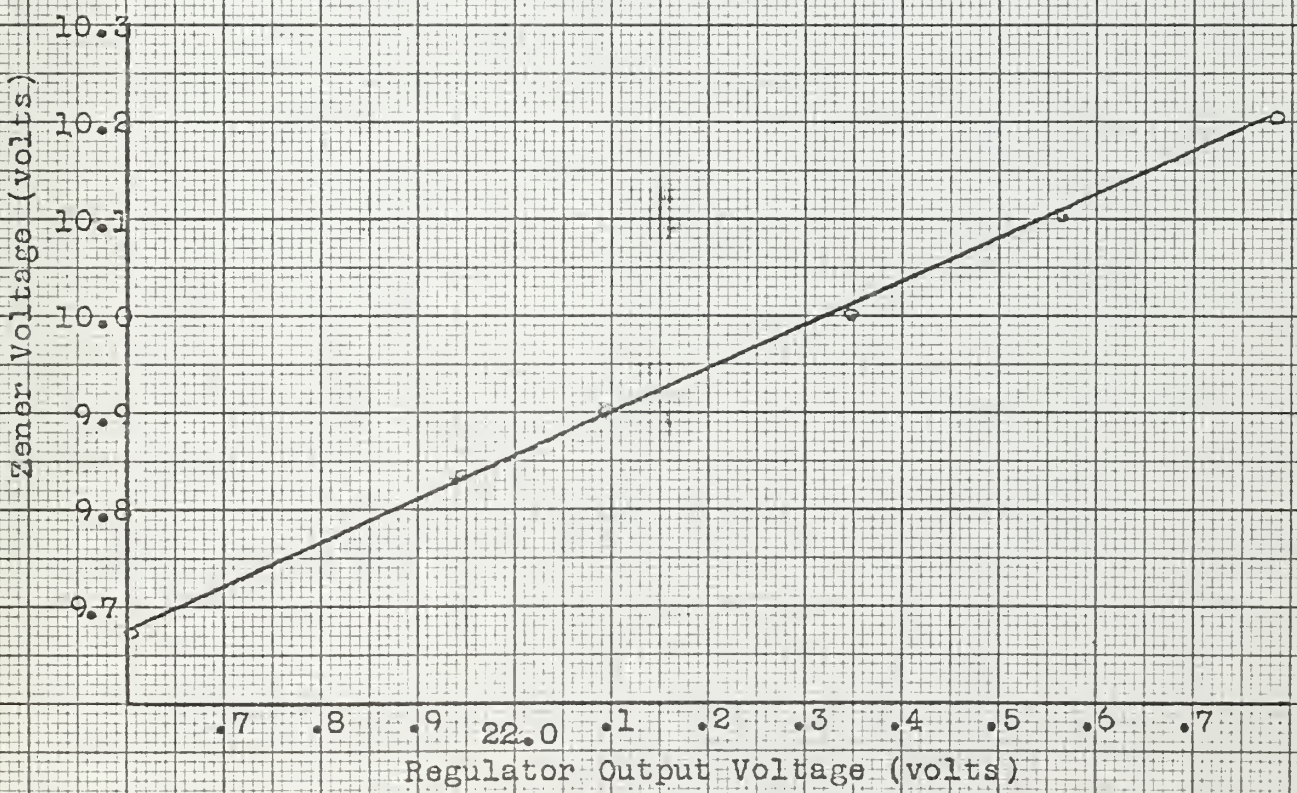


Figure 7(b) - Regulator Output Voltage as a Function of Zener (Reference) Voltage.

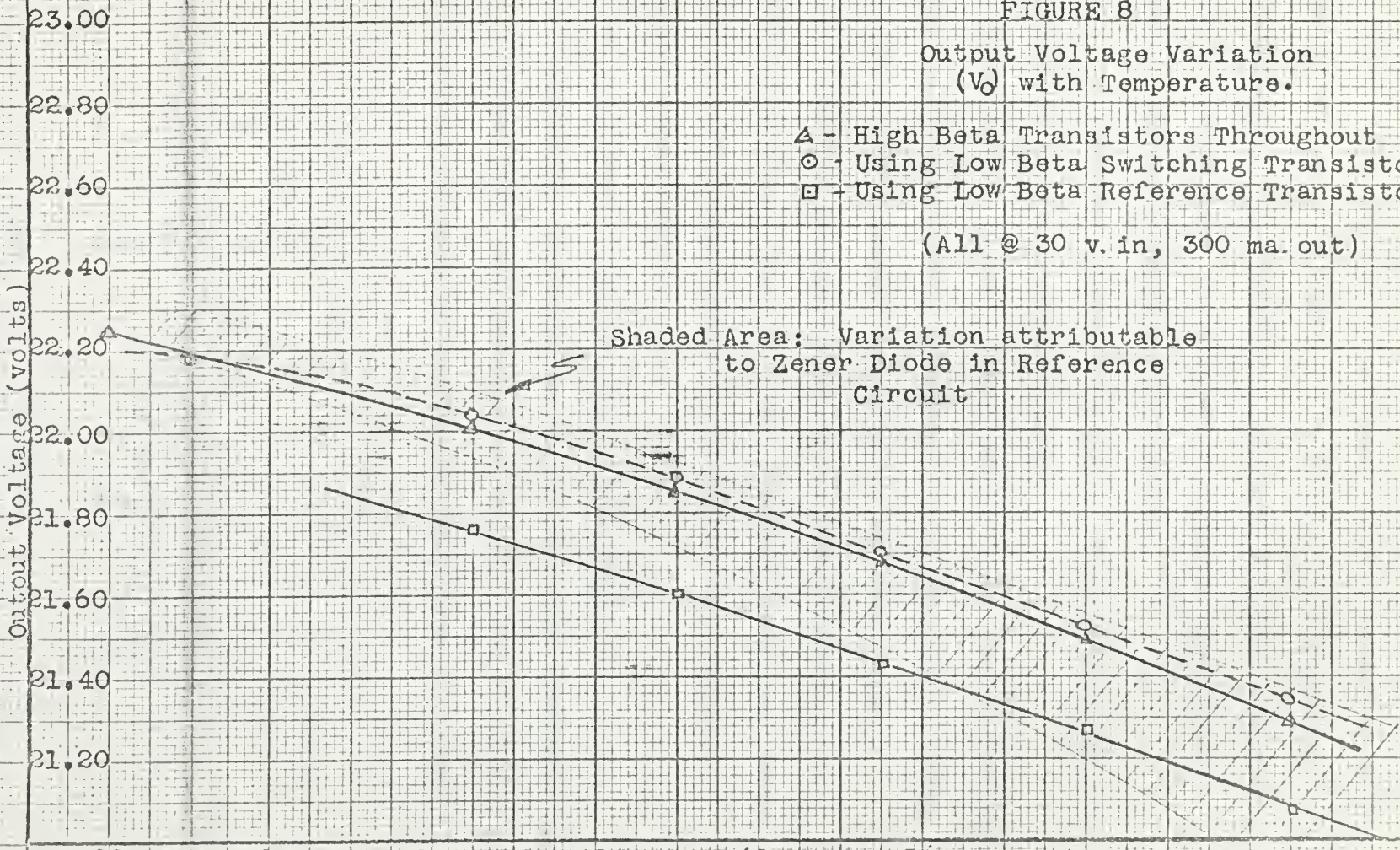


FIGURE 8

Output Voltage Variation (V_o) with Temperature.

△ - High Beta Transistors Throughout
⊙ - Using Low Beta Switching Transistors
⊡ - Using Low Beta Reference Transistor

(All @ 30 v. in, 300 ma. out)

SWITCHING CIRCUIT

Clean, hard, predictable switching action is a prerequisite of output voltage stabilization. To provide these desirable characteristics, three parameters governing switching operation had to be investigated, and minimum requirements equalled or exceeded. The requirements are listed below:

1) An invariant reference voltage be maintained within $\pm$ 10 millivolts at the first switching stage.

2) An overall current multiplication factor exist sufficient to cleanly switch the power transistor on and off without loading the voltage divider circuit supplying the switching signal.

3) Switching transient times be of constant duration, unaffected by regulator input or output conditions, and fast enough so that time lag between switching initiation and completion do not affect output voltage.

Two conditions of the switching circuit exist:

Switch on: Configuration to turn the power transistor on.

Switch off: Configuration to turn the power transistor off.

Both conditions are shown in Figures (9a) below, and (9b) Page 15.

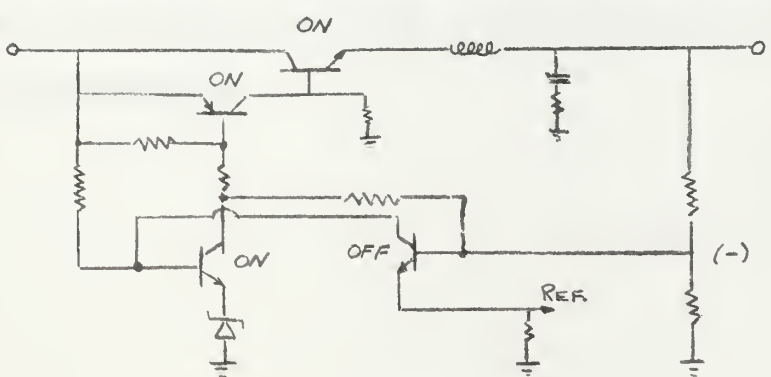


Figure (9a) - Switch on

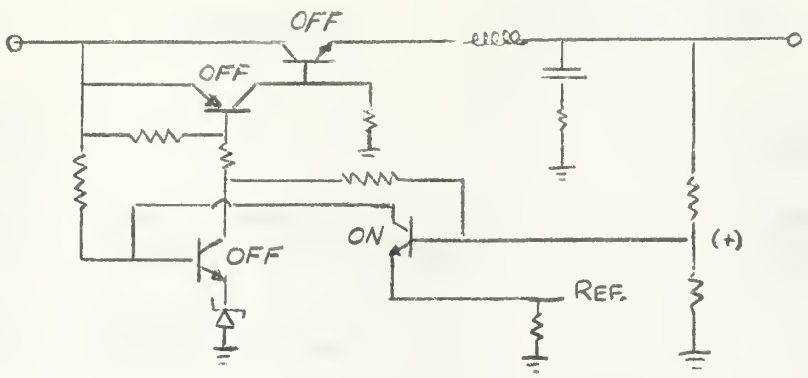


Figure (9b) - Switch off

The three requirements for predictable switching were satisfied using the following methods:

1) An invariant reference voltage was obtained at the emmitter of the initial switching transistor by use of the voltage reference circuit described in Part I.

2) An overall current multiplication factor sufficient to amplify the switching signal without loading the detection circuit was obtained using high beta transistors (npn - 2N-1711, Beta approx 150, and pnp-2N1132, Beta approx 60), although investigations were carried out using lower Beta transistors (npn-2N697, Beta of 90; npn-2N696, Beta of approx. 45) with results as shown in Table II the next page (Page 16).

	Regulation	Efficiency @ 30 Volts, .3 amp	Comments
2N335 B=30	-	78%	No switching @HiLoad, Lo input
2N696 B=45	Load: 5mv@30V Voltage: 92 mv@.4Amp	79%	Rounded switching Signal
2N697 B=90	Load: 10 mv Voltage: 15 mv	82%	-
2N1711 B 120	Load: 10mv Voltage: 14mv	83%	-

Table II - Comparison of operation using Low Beta Transistors.

Although the 2N696 transistor apparently has better load regulation characteristics at constant voltage input, voltage regulation and efficiency are well below optimum. Switching action in general was not as hard using the low Beta transistors, switching corners were not square and high harmonics not present, indicating use of a non-saturating load line in the first two switching stages.

3) Switching transient times were made immune to input voltage by operation of the switching transistors in a saturated condition, and use of high beta transistors. Switching transient times were made small by use of a regenerative, or positive feedback, path from second to initial switching stages. All stages after the initial were switched between saturated and non-conducting configurations in times governed by the limitations of the transistors and switching signals from the initial switching stage.

A step by step commentary following the switching signal is the best description of the switching circuit. All references to components refer to Figure (2).

1) Potential drops at the base of the first stage switching transistor, shutting it off to a nonconducting configuration. Because the em itter voltage is constant (controlled by the reference stage), collector potential rises along a load line as determined by R_5.

2) A switching signal appears across the load resistor R_5, which is common to the input (base) of the second stage. An increase in V_c of the first stage results in an increase of V_b second stage. Because V_e second stage is controlled by a zener diode, it is relatively stable at 12 Volts, and an increase of V_b results in an increase of V_{b_e} which turns the second stage on to a saturated conducting configuration, the potential at the collector approaching V_e. Current through the second stage is limited by the value of R_6, and must be sufficient to switch the third stage. Feed back current used to assist in biasing the initial stage through R_7 is cut off, first stage switching action is driven harder into nonconduction, and regeneration causes the first and second stages to switch even harder in their respective directions. The relative effect of bias of the first stage by R_7 is determined by the percent of current it supplies, as given by the ratio

$$\% \text{ Current} = 1 - \frac{R_7}{R_9 + R_6 + R_7}$$

The net effect of the variation of the feedback resistor with respect to switching frequency and ripple voltage dead zone is shown in Figures (11) and (12).

3) A square wave current signal appearing at the base to the third stage is amplified by the third stage sufficiently to provide the power transistor T_1 with sufficient power for switching. R_4 is a small resistor of approximately 150 ohms to prevent temperature leakage of the third stage.

4) The power transistor is switched on by current supplied by the third stage.

The opposite sequence of operation occurs to turn the power transistor off, as shown diagramatically in Figure (9b). Switching signals at various points in the circuit are shown in Figure (10) and Appendix 1. Alternate circuits were attempted using fewer components, but none proved as efficient or reliable due to a weakness in obtaining a hard reference voltage to initiate switching.

Efficiency losses are discussed in the last section.

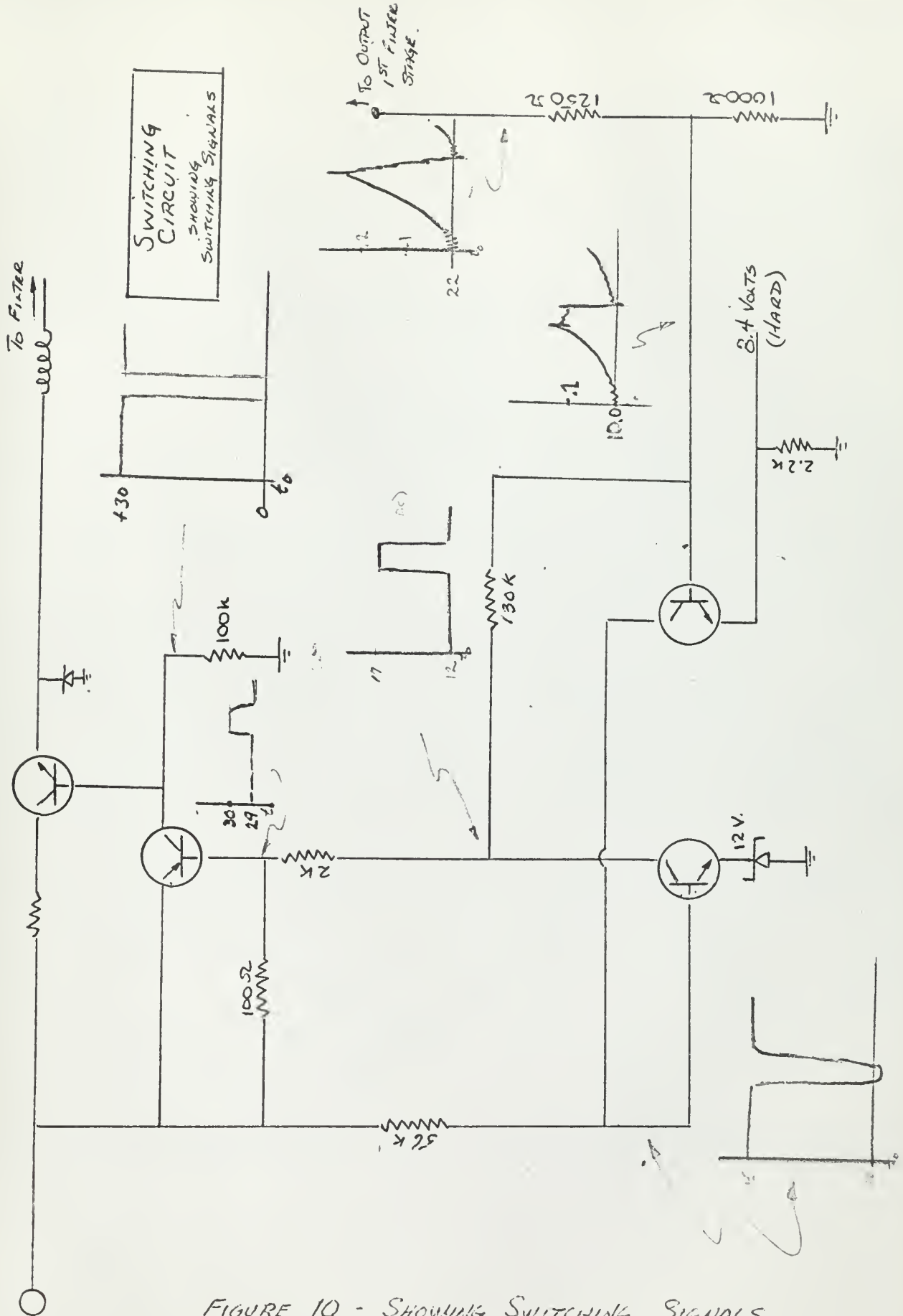


FIGURE 10 - SHOWING SWITCHING SIGNALS

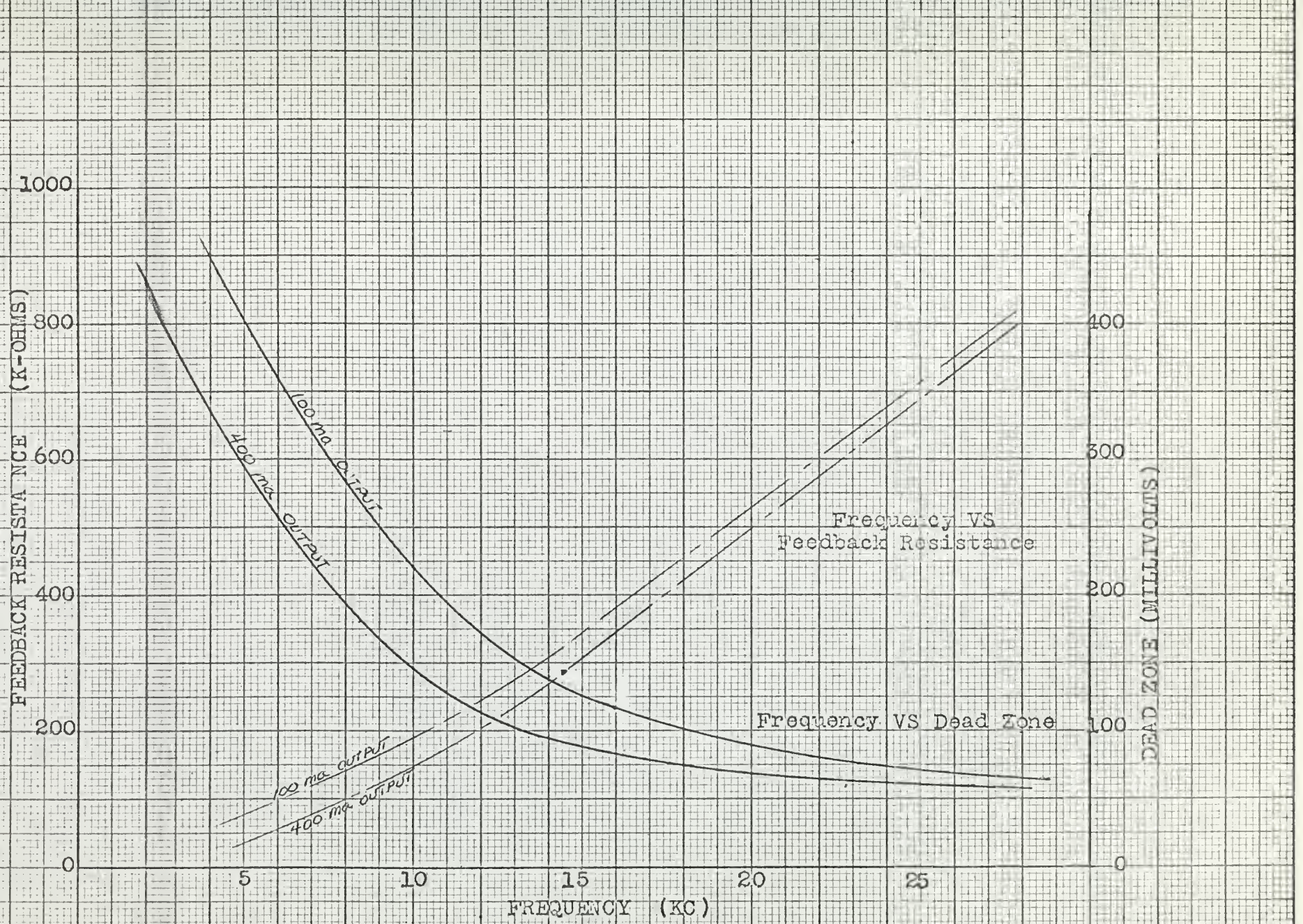


Figure 11 - Variation of Frequency with Feedback Resistance and Dead Zone

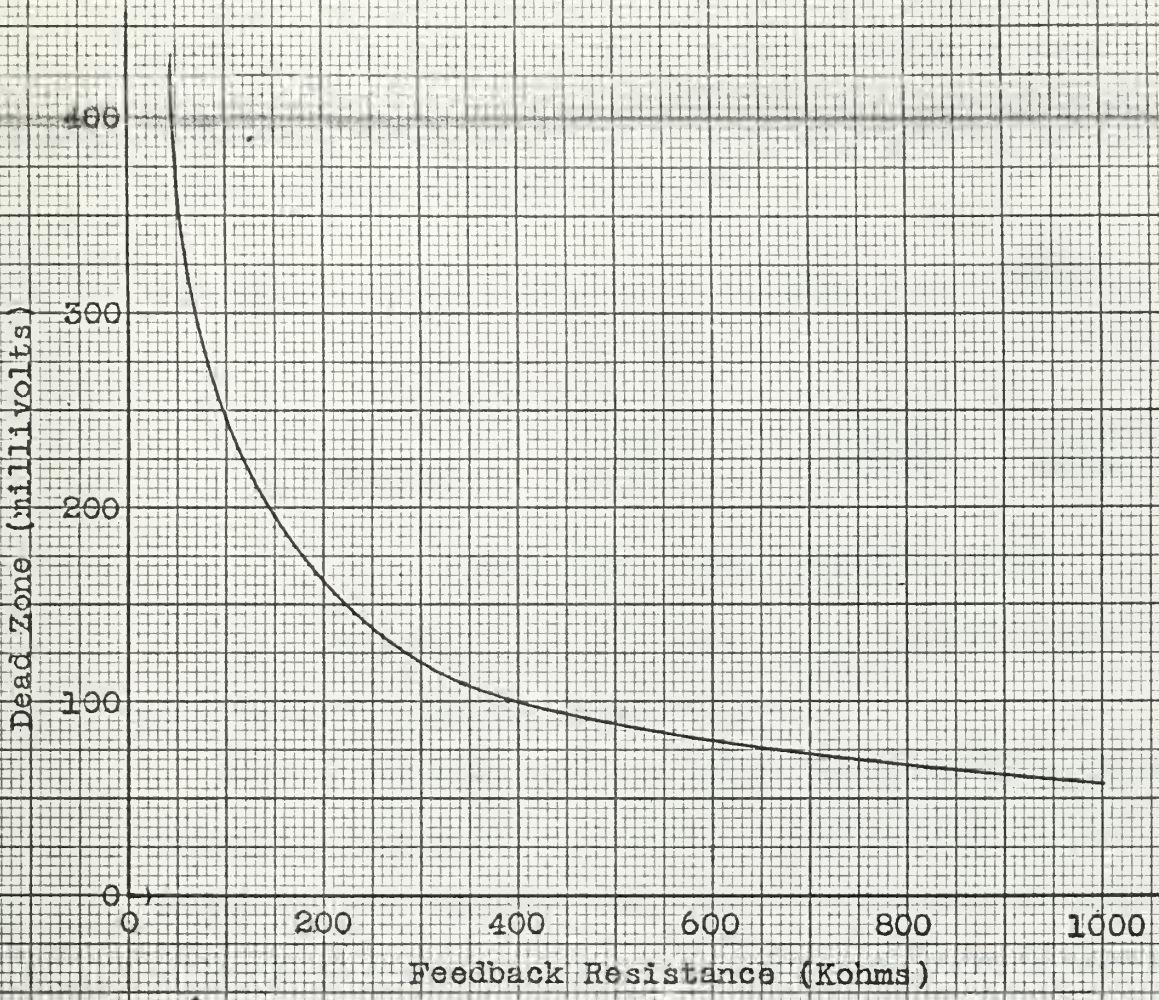


Figure 12 - Effect of Variation of Feedback Resistance on Dead Zone at 100 ma Load and 30 Volts Input.

The power conduction circuit is composed of the main power switching transistor T_1, an energy storage inductor L_1 wound on a powdered iron core, and capacitor C_1, and associated filter components as shown in Figure (13) below:

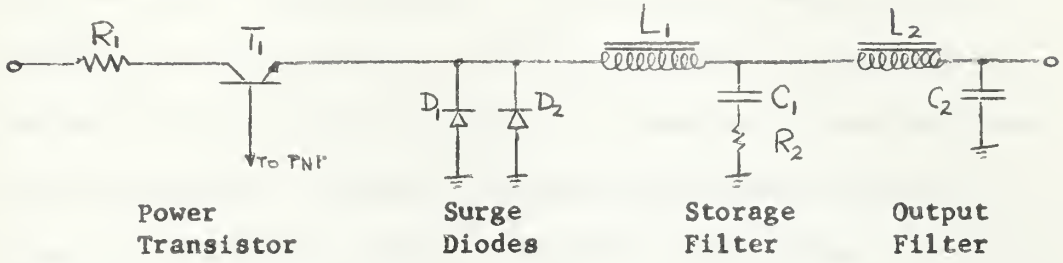


Figure (13): Power Conduction Circuit

If a theoretically perfect power circuit were to be made it would have the following characteristics:

A switch would be closed to conduct current at input potential to load the storage coil. This switch would be 100% efficient, having infinite resistance when open, zero resistance when closed, infinitesimal switching times, and requiring no power to complete switching. The output square wave would then be absolutely square in both current and voltage, and would be used to load a coil of infinite Q and zero core loss. No high frequency harmonics would be coupled to any other part of the circuit and output ripple voltage would be infinitely small.

Obviously such a circuit is unattainable so the following compromises are made.

1) Accept present power transistor switching speed and power dissipation specifications and design for maximum efficiency and

proper operation around this component.

2) Accept an output ripple voltage of between 100 - 200 millivolts.

There is no switching power transistor presently available that can be switched instantaneously. The transistor used (Lockheed #1960893) switched from off to full on in about 2 u sec. Power, absorbed during the switching cycle, is required to turn the transistor on and off. Although I^2R losses also occur across the transistor, switching losses, as measured by the shaded area shown below in Figure (14), are instrumental in determining the efficiency of the overall regulator. Variation of efficiency with frequency is shown in Figure (18).

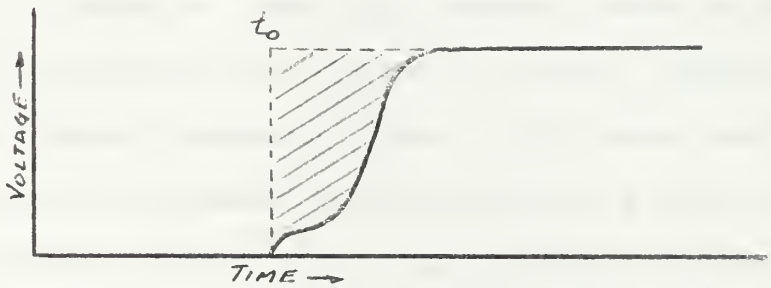


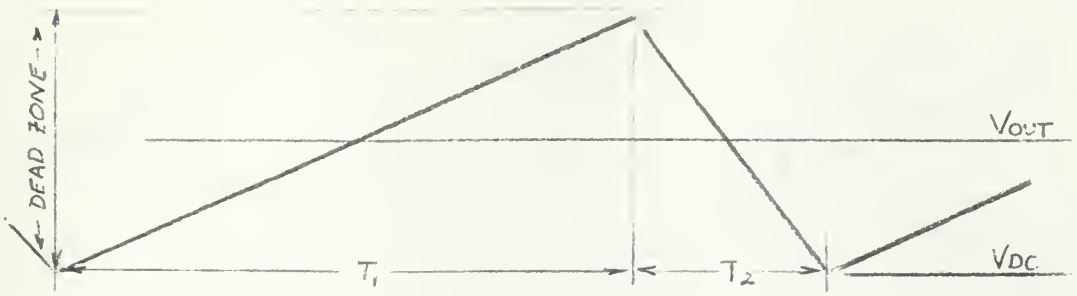
Figure (14) Relative power to complete switching.

Unfortunately, a square wave signal input of adequate power to measure the rise time explicitly was not obtainable, therefore all data taken is reported relative to the most efficiently operating power transistor obtainable (Lockheed #1960893). A relative comparison between a fast and slow transistor is shown in Plate 2 of Appendix 1. It is worthy to note that the slow transistor ran hot at full power, whereas the fast one did not. Both had the same power rating.

Voltage drop present across the power transistor in the circuit is instrumental in determining minimum input voltage required to load the inductance coil. As input voltage decreases, frequency decreases as shown in Figure (3). Note that 24 volts is not sufficient to maintain switching frequency above 1Kc, resulting in a loss in regulation control as the power transistor remains a closed switch, and the system acts as a series regulator

Surge diodes D_1 & D_2 in the emmitter circuit of the power transistor allow a spike of current to flow from ground into the storage coil L_1 when the power transistor switches off, as shown in Plate 4 of Appendix 1. Coupling of this high frequency spike into various parts of the circuit results in extreme loss of efficiency and erratic operation. Other high frequency harmonics present in the switching signals can be traced to the power transistor, probably to frequency components present in the square wave output. These are partially attenuated by the addition of a small resistor R_1 in the collector circuit of the power transistor which damps the high frequencies to an acceptable level. Other slower power transistors did not require Resistor R_1 to damp out high frequencies.

Assuming that a filter is used on the output to attenuate the switching signal, a finite ripple voltage of 100 - 200 mv to start and stop switching is realistic. The circuit is therefore limited in switching frequency by the rate of charge and discharge of the storage inductor - capacitor combination within the "dead zone" of the ripple voltage. Graphically the circuit has a switching frequency related to dead zone in the following manner:



T_1 Time required to load inductor - capacitor (L_1 - C_1) to V_{out} + ½ D

T_2 Time required to discharge inductor - capacitor (L_1 - C_1) to V_{out} - ½ D

Figure (15) - Relation of frequency to dead zone.

Peak repple voltage values are relative to the final filtered DC level for an integrating filter Thus, for a triangular switching signal shown the output voltage is quiescent at one-half the height of the dead zone. Notice that a distorted ripple voltage within this dead zone of any non-symmetrical shape will not yield V_{out} as its integral. Therefore, to prevent output voltage variation, the switching signal may vary only in the manner such that its integral remains constant. A small resistor of the order of one ohm (R_2) prevents ripple signal waveform variation with load by effectively acting as a phase lead network for the switching signal to the initial switching stage. This resistor is in fact one of the keys to reliable switching action, as shown in the comparison of ripple voltage waveforms at high output power and low input voltage of Figure (16) on Page 26 and in Plate 1 of Appendix 1.

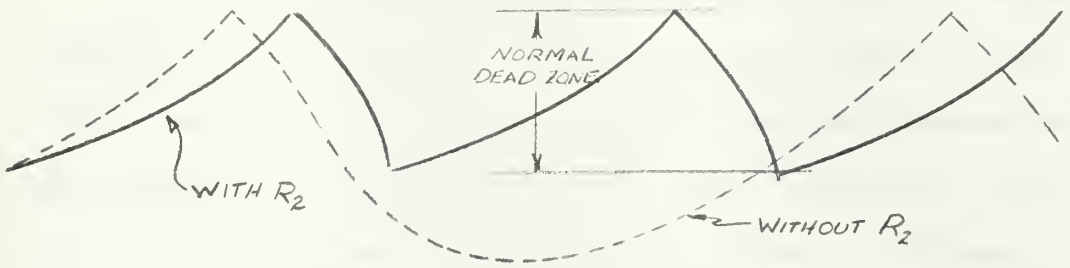


Figure (16) - Comparison of ripple voltage wave forms with and without R_2 at 400 ma and 25v input.

Integration of the distorted ripple signal yields some value less than the quiescent DC value of an integrated sawtooth wave, thus affecting output regulation by decreasing the level of filtered DC output voltage at the first switching stage. A qualitative frequency analysis, with and without the resistor, is shown in Figure (17) below.

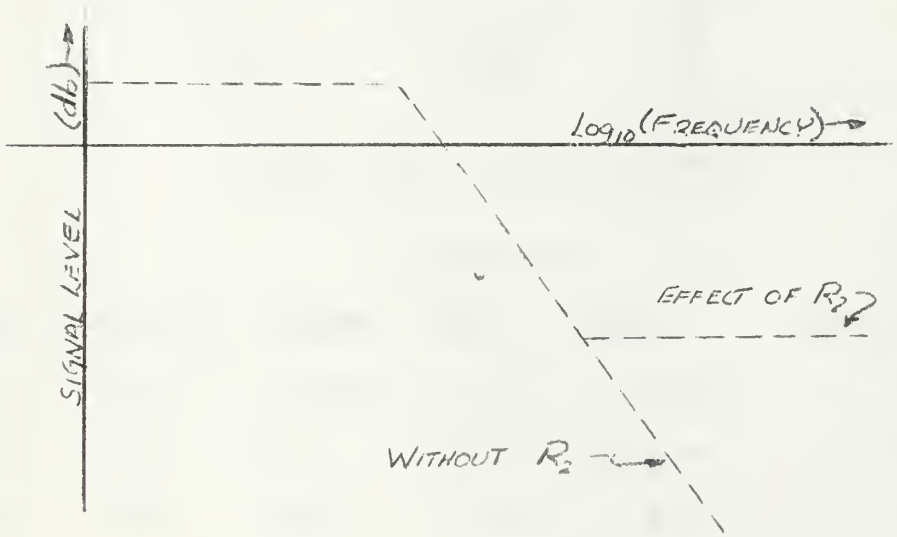


Figure (17) - Frequency attenuation of ripple signal with and without resistor R_2.

DISCUSSION OF EFFICIENCY

A basic purpose for the use of a variable duty cycle voltage regulator is the efficiency of operation at high loads; a good series regulator can achieve at least the same ouput regulation as the specified .1%, but lacks efficiency by the nature of its operation. A discussion of the sources of loss of efficiency of the variable duty cycle regulator is presented to show where future improvements need to be made to achieve even greater superiority by use of improved circuit components. Table (III) is presented in conjunction with Figure (18) to show where power is absorbed in the circuit.

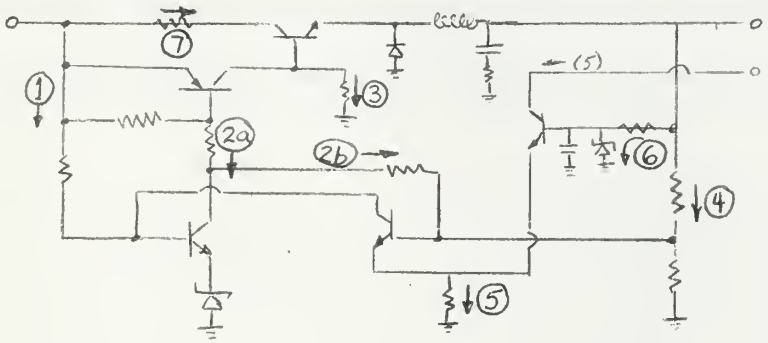


Figure (18) - Efficiency Losses.

Table (III)

PATH	VOLTAGE	CURRENT	TIME	POWER	% of 8.8 watts
1	16.5	.295 ma	.75	3.65 mw	.04
2a	15	7.5	.75	112.5 mw	.95
2b	30	.2ma	.25	3.0 mw	.04
3	30	.3ma	.75	6.8 mw	.08
4	22.0	1ma	-	22.0 mw	.27
5	9.4	4.27ma	-	40 mw	.49
6	12.0	1ma	-	12 mw	.18
7	R - 1ohm	.4 amp	(Avg)	160	1.90
			Total Switching Power (%)		3.95

From Table (III) it can be seen that the switching and voltage reference circuits account for only about 1/3 of the power loss of the circuit. The other 2/3 is used in the combination of the power transistor, inductor, or capacitor. Because the power transistor runs warm during high load conditions it appears to be the main source of power loss in the circuit. Whether these losses are a result of switching or I^2R losses across the transistor is not apparent, but Figure (18) shows the overall efficiency as a function of frequency thus providing a qualitative look at the increase of power required as switching is increased. Because transistors are still undergoing development, it is possible that future power transistors may have extremely high efficiencies, thus allowing overall regulator efficiencies of greater than 95%.

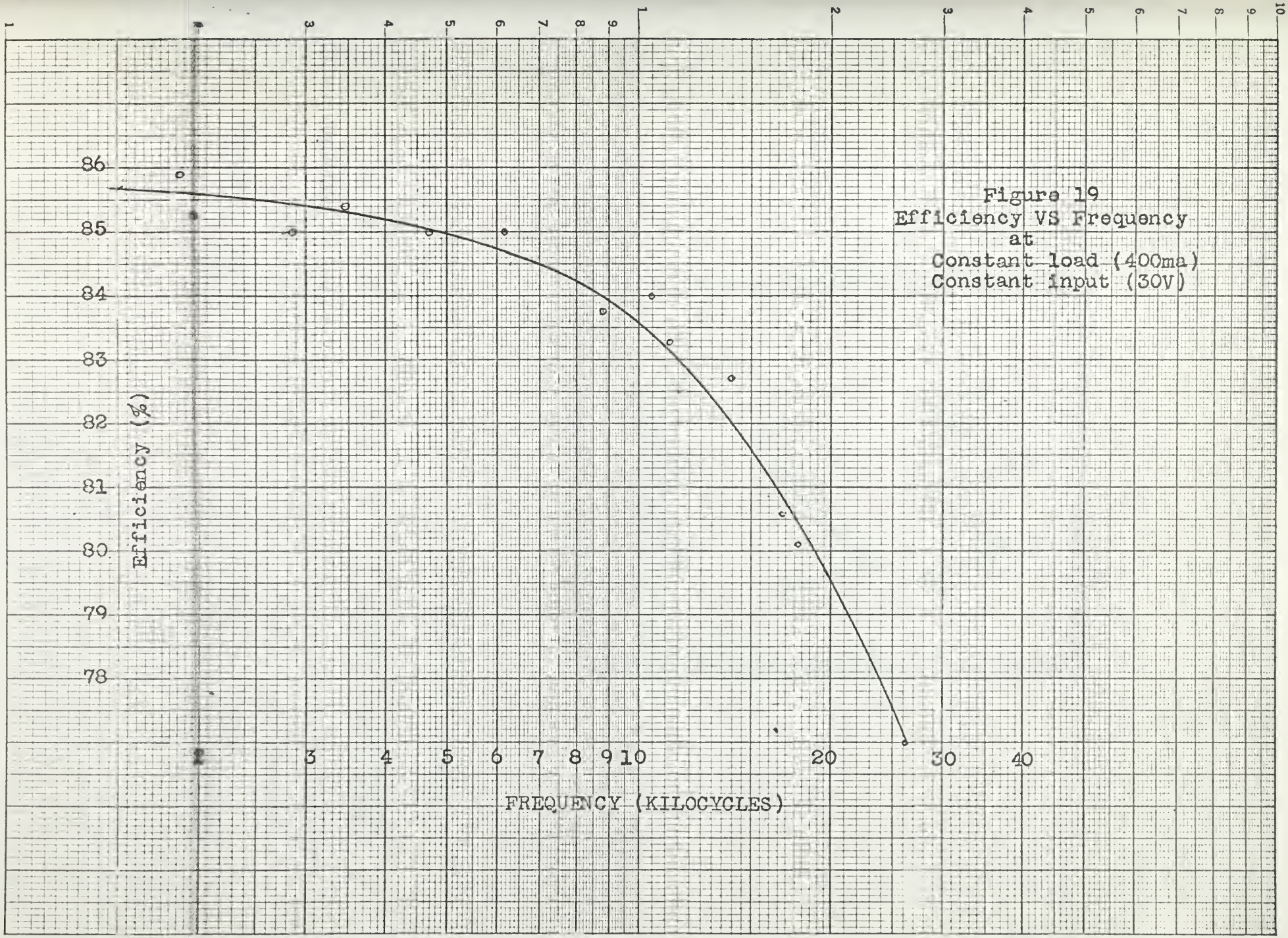


Figure 19
Efficiency VS Frequency
at
Constant load (400ma)
Constant input (30V)

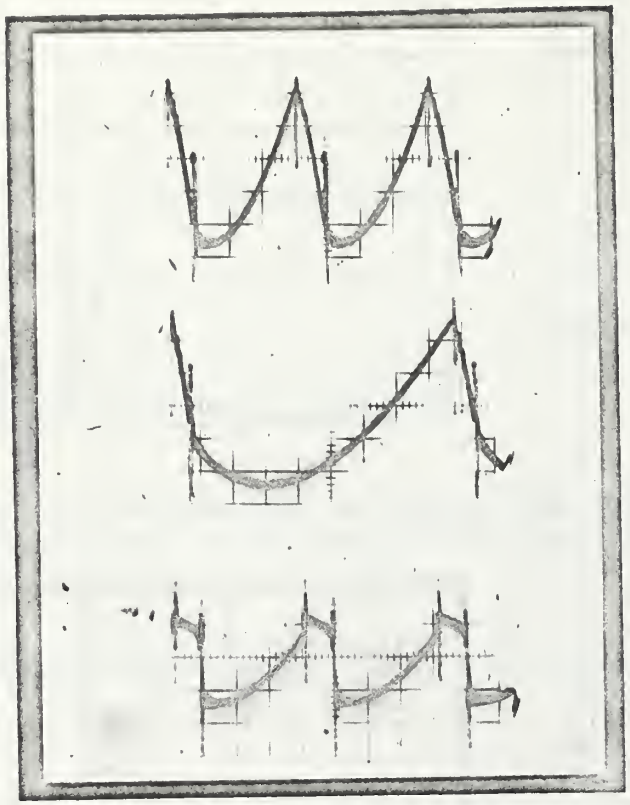


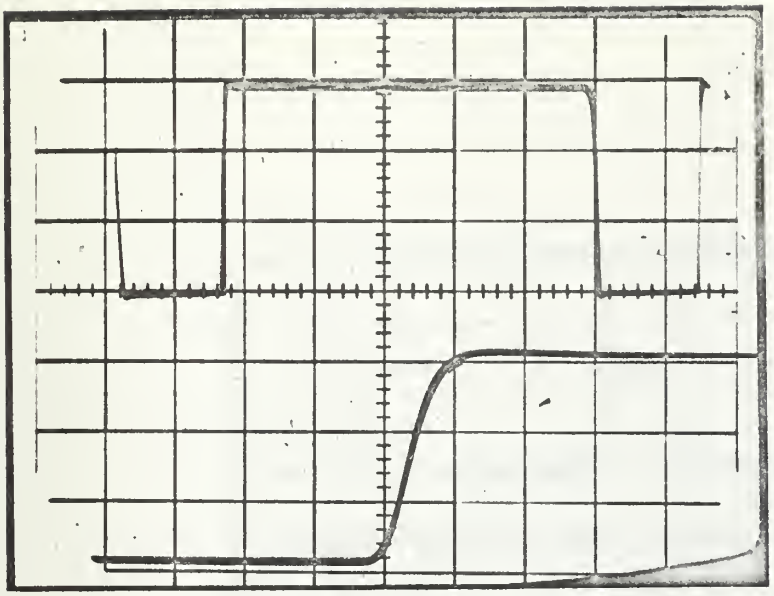
Plate I

Comparison of switching signals with and without resistor R_2.

Top: Ripple signal with R_2 installed, before filtering.
Vertical: 50 mv/cm. Horiz: 50 usec/cm.

Middle: Ripple signal without R_2 installed, showing ringing and enlarged dead zone.
Vertical: 50 mv/cm. Horiz: 50 usec/cm.

Bottom: Signal at input of first stage switching.
Vertical: 50 mv/cm. Horiz: 50 usec/cm.



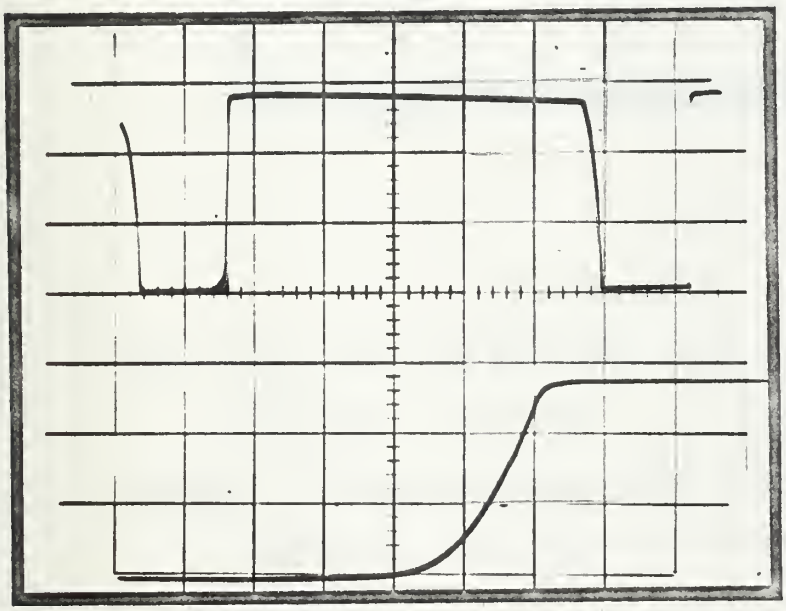
Vertical: 10 v/cm

Horizontal: 20 usec/cm

Vertical: 10 v/cm

Horizontal: 1 usec/cm

Plate II(a) - Switching of fast, low I^2R loss power transistor, 30 v. input, 300 ma. load.



Vertical: 10 v/cm

Horizontal: 20 usec/cm

Vertical: 1o v/cm

Horizontal: 1 usec/cm

Plate II(b) - Switching of slow, large I^2R loss power transistor, 30 v. input, 400 ma. load.

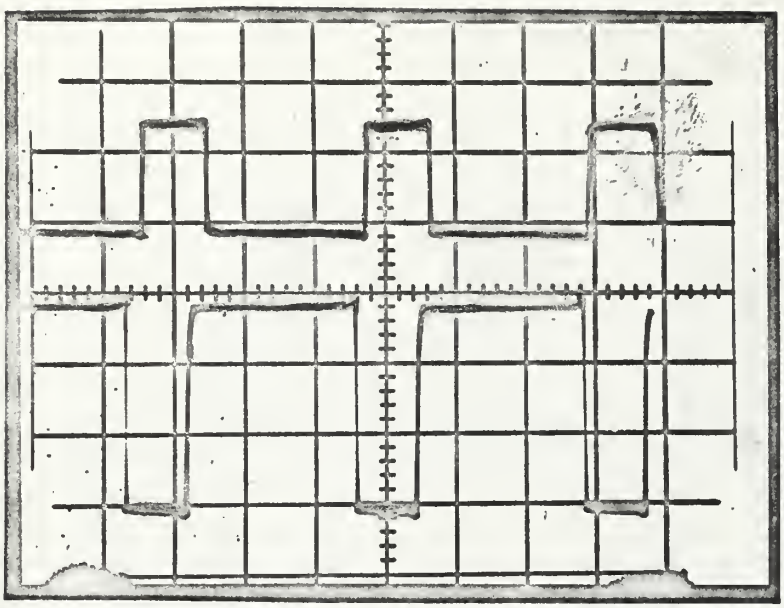


Plate III

Top: - I_b across 2K resistor into PNP switching stage.

Vertical: 10 v/cm Horizontal: 50 usec/cm

Bottom: Voltage driving power transistor (V_b).

Vertical: 10 v/cm Horizontal: 50 usec/cm

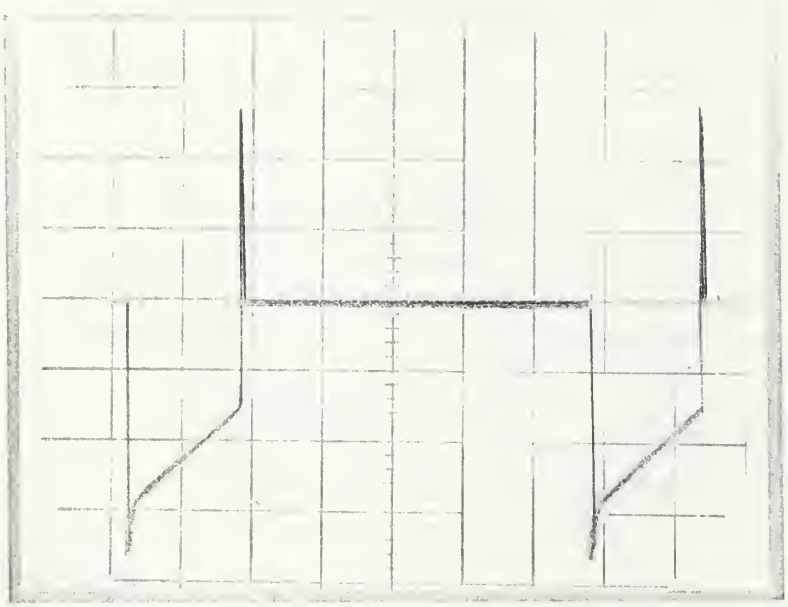


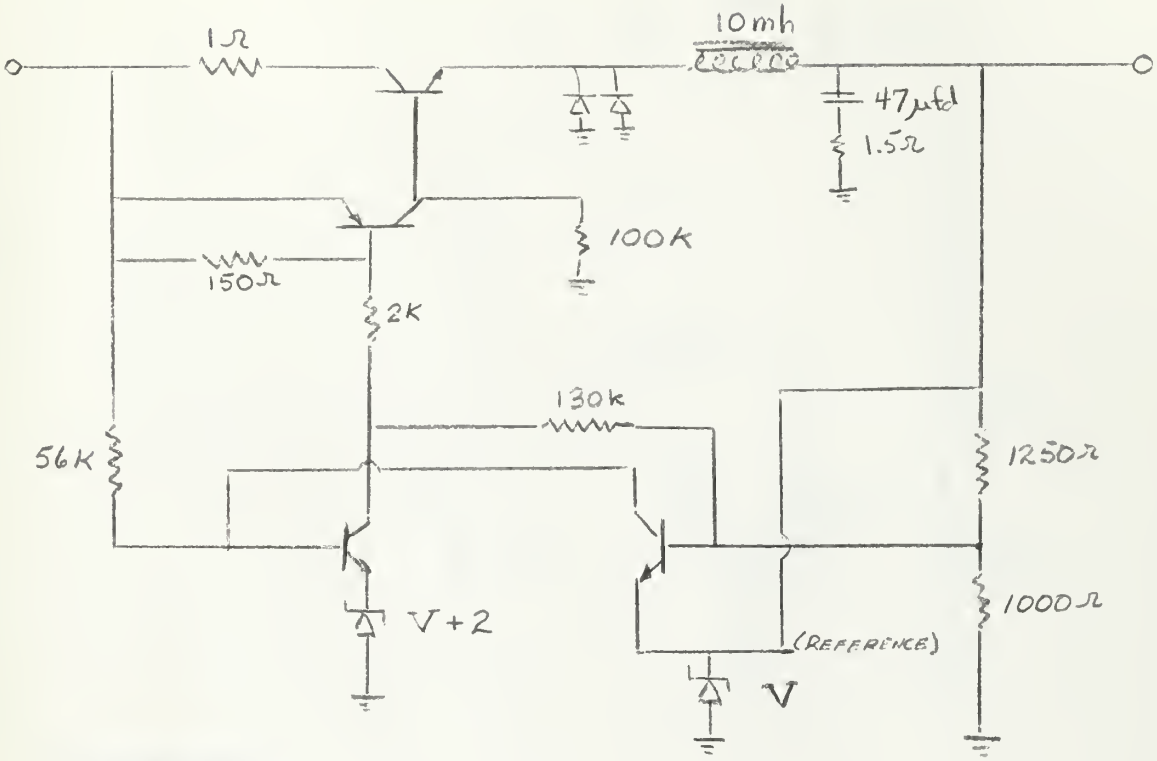
Plate IV

Voltage across a one ohm resistor in diode circuit, showing action of surge diodes and current spike.

Vertical: 100mc/cm Horizontal: 20 usec/cm

APPENDIX 2

ALTERNATE CIRCUITS

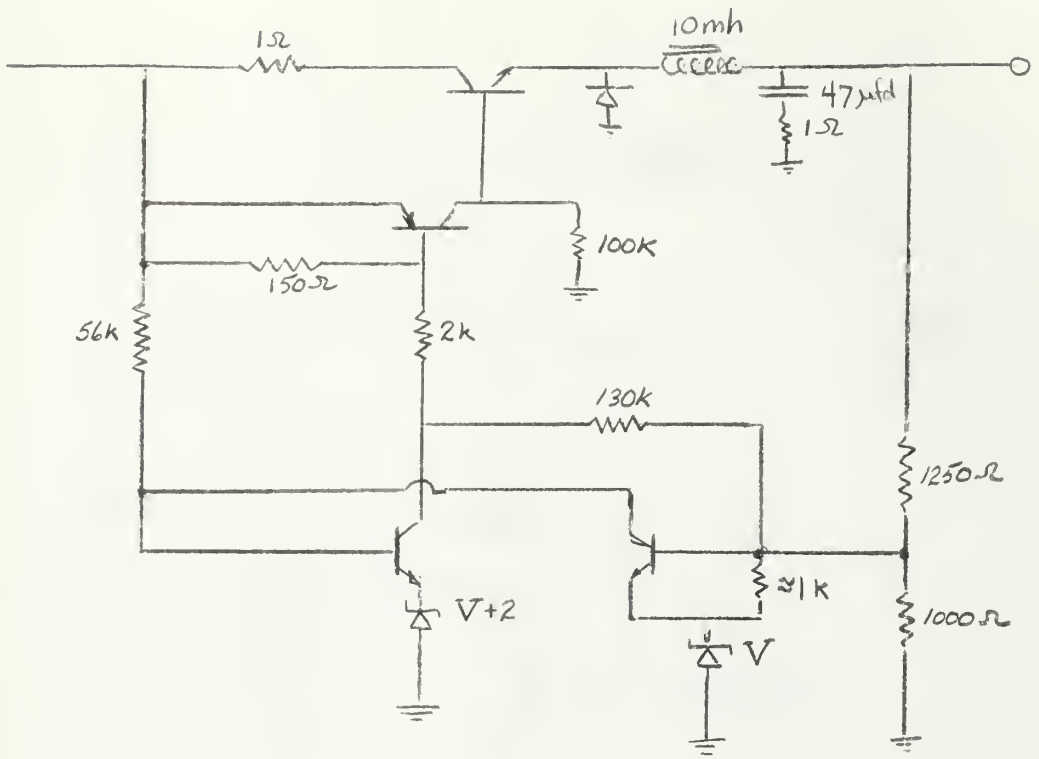


DISADVANTAGES

1) "V" denomination zener diode is in the emmitter of the switching circuit and is therefore subject to more than the 1 ma at which its temperature rating is given.

2) Reproducibility of circuit characteristics is poor due to the effect of variation of the knee of the zener diode, and the amplification of this variance.

3) Reference voltage "floats" excessively due to the large variation of current through the zener diode during switch on and switch off cycles.



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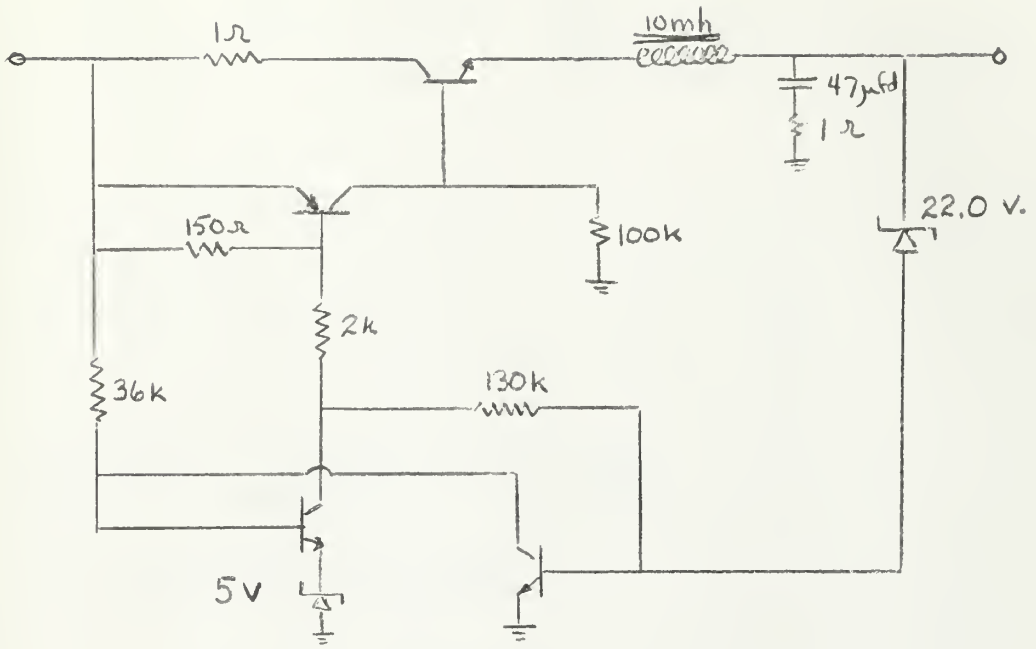


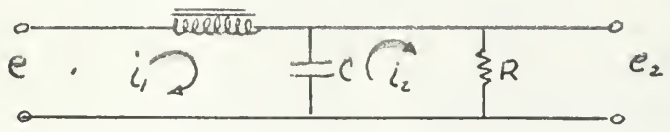
DISADVANTAGES

1) Cost of obtaining a 1% 22.0 volt zener diode is excessive.

2) Matched zener diode pair (22.0 volts) is required to satisfy temperature coefficient specifications.

APPENDIX III

FILTER CIRCUIT:



$$\tilde{e} = LS\tilde{i}_1 + \frac{1}{CS}(\tilde{i}_1 - \tilde{i}_2) \tag{1}$$

$$0 = R\tilde{i}_2 + \frac{1}{CS}(\tilde{i}_2 - \tilde{i}_1) \tag{2}$$

$$\therefore \tilde{i}_1 = (CSR + 1)\tilde{i}_2 \tag{3}$$

$$\tilde{e} = \left[(LS + \frac{1}{CS})(CSR + 1) - \frac{1}{CS}\right]\tilde{i}_2$$

$$= (CLRS^2 + LS + R)\tilde{i}_2$$

$$\therefore \tilde{i}_2 = \frac{\tilde{e}/CLR}{(S^2 + S/CR + 1/LC)} \tag{4}$$

$$\therefore \tilde{e}_2 = i_2 R = \left\{\frac{1/CL}{S^2 + S/CR + 1/LC}\right\}\tilde{e} \tag{5}$$

REAL ROOTS WHEN $L = 4CR^2$

FOR $L = 10mh$, $C = 50\mu fd$, $R = 7.5$ OHMS.

SOLUTION OF EQN (5) IS OF THE FORM

$$\tilde{e} = \frac{kE}{S(S+\alpha+j\beta)(S+\alpha-j\beta)} \tag{6}$$

$$e \cong 1 - \epsilon^{-\alpha t}\sin(\omega t + \phi) \tag{7}$$

WHICH IS A DAMPED OSCILLATION.

where $\alpha = 1/2CR \approx 200$ AT 400 ma LOAD

$\beta = (1/4C^2R^2 - 1/LC)^{1/2}$

$\approx 1,414$ AT 400 ma LOAD

$\phi = \tan^{-1} \beta/\alpha \approx 82°$

$\omega = (\alpha^2 + \beta^2)^{1/2} \approx 1,500 \frac{RAD}{SEC}$

$\approx 86 \times 10^3 \frac{deg}{sec.}$

using $L = 10$ mh
$C = 50$ μfd
$R = 55$ OHMS
} SIMULATING FULL LOAD.

BIBLIOGRAPHY

1. G. N. Patchett, Automatic Voltage Regulators and Stabilizers, Sir Isaac Pitman & Sons, Ltd., 1954.

2. Fairchild Corporation, Data Sheets SL-4, 5, 7, March 1959